

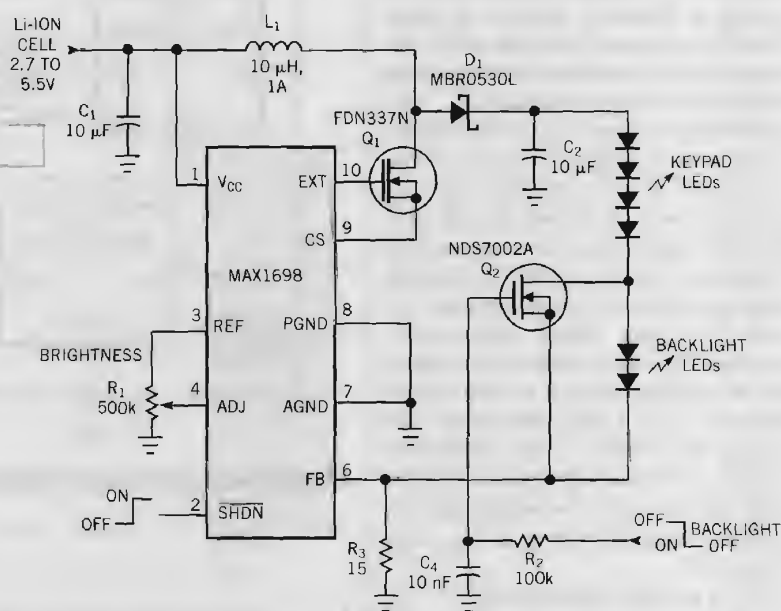
Edited by Bill Travis and Anne Watson Swager

## Single FET controls LED array

Len Sherman, Maxim Integrated Products, Sunnyvale, CA

**W**HITE-LED BACKLIGHTS are gaining acceptance because they offer higher reliability and simpler drive circuitry than backlights based on CCFL (cold-cathode-fluorescent-lamp) and EL (electroluminescent) technology. As a result, white-LED backlights are increasingly common in PDAs (personal digital assistants), cell phones, digital cameras, and other portable devices. A design in which the display requires backlighting for extended periods needs an efficient circuit that drives the LEDs with a controlled current and eliminates the wasted power associated with current-limiting resistors. **Figure 1** shows a switch-mode boost design that regulates current instead of voltage. Because all the LEDs are connected in series, they all receive the same current without the need for ballasting resistors. Identical currents help achieve uniform intensity. And, because the output current is low (20 mA in this case), the output-filter capacitance,  $C_2$ , can be smaller than for a load consisting of parallel-connected LEDs.

**Figure 1**



When this circuit turns off the backlight LEDs, the keypad LEDs remain on with no change in intensity.

The circuit's 90% conversion efficiency offers a distinct power-saving advantage over resistor-limited and linearly regulated designs. It might appear that a series-LED connection is unsuitable for applications in which some (but not all) LEDs must be off. A cell phone, for example, sometimes needs that capability for occasions when the display is off but the keypad remains lit. Or, a PDA might need to play a sound file while maintaining illumination in the buttons but not the display. In the circuit of **Figure 1**, switching off individual LEDs or groups of LEDs is not a problem, even with series drive. Applying a logic-high level to the gate of a simple MOSFET switch,  $Q_2$ , turns off a subset of LEDs by shunting

their current. The remaining LEDs (for the keypad, for example) remain on, and their intensity remains constant because  $IC_1$  regulates their current, by sensing the voltage across  $R_2$  (300 mV at full brightness). When the circuit turns the LEDs on and off, the  $R_2$ - $C_4$  network at the gate of  $Q_1$  slows the load changes sufficiently to prevent transients in the LED drive current. Other features include adjustable intensity via the ADJ pin and full shut-down via the SHDN pin.

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Single FET controls LED array .....	131
Circuit protects battery from overdischarge .....	132
Two diodes change demagnetization-signal polarity .....	134
Simple scheme keeps current drain constant .....	138
RS-232/485 converter has automatic flow control .....	140
Circuit provides accurate RTD measurements .....	142

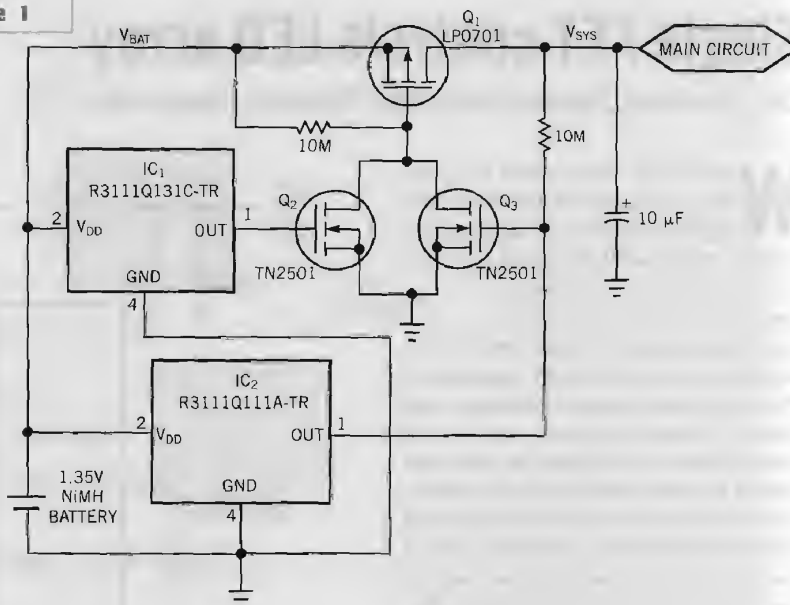
# Circuit protects battery from overdischarge

Martin Wuzik, Implex AG Hearing Technology, Ismaning, Germany

**A**LL ELECTRONIC CIRCUITS powered by a battery discharge the battery. In some applications, it is undesirable to overdischarge the battery, because it could irreversibly reduce the battery's capacity and the number of discharge/charge cycles. The circuit in **Figure 1** protects a single NiMH (nickel-metal-hydride) cell by disconnecting the load from the battery. **Figure 2** shows the output voltage,  $V_{SYS}$ , versus the input voltage,  $V_{BAT}$ . For this NiMH battery, the switching points are 1.1 and 1.3V. If the battery discharges and  $V_{BAT}$  drops below 1.1V,  $Q_1$  switches off, and the node Main Circuit disconnects from the battery. In that case, the battery's only load is the pair of voltage detectors  $IC_1$  and  $IC_2$  from Ricoh (www.ricoh-usa.com). The load current of one detector is typically 800 nA, so the battery drain is 1.6  $\mu$ A. The user must now charge the battery. Once the battery charges and the voltage reaches 1.3V, the load reconnects to the battery and remains connected as long as  $V_{BAT}$  stays above 1.1V.

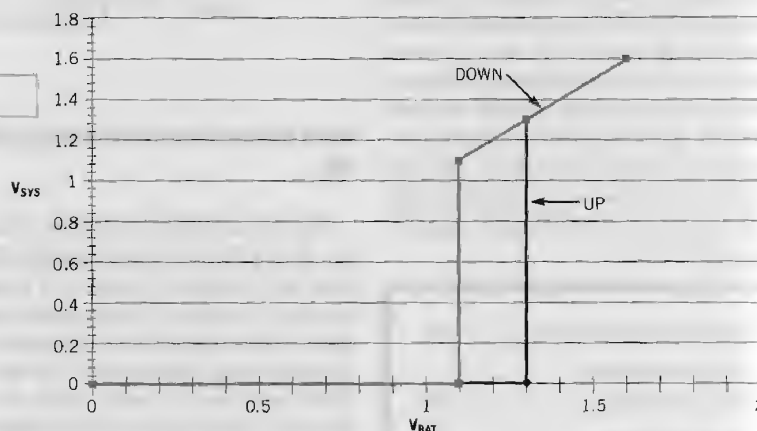
$IC_1$  is a voltage detector with a 1.3V setpoint and a push-pull output.  $IC_2$  has a 1.1V setpoint. An important difference between the two detectors is that  $IC_2$  has an open-drain output. If the battery voltage drops but remains within the 1.1 to 1.3V range,  $IC_1$ 's output is low, and  $Q_2$  switches off.  $Q_3$  switches on because  $IC_2$ 's output is still in the high-impedance state. If  $V_{BAT}$  drops below 1.1V,  $IC_2$ 's output switches low,  $Q_3$  turns off, and, as a result,  $Q_1$  also switches off. As soon as  $V_{BAT}$  drops below 1.1V, the load disconnects from the battery. The load reconnects to the battery only when the battery charges to a voltage higher than 1.3V. At voltages of 1.1 to 1.3V,  $IC_2$  cannot switch on  $Q_3$  because the IC's output is an open-drain type and  $V_{SYS}$  is low.  $IC_1$ 's output must assume a high state to switch on  $Q_2$  and to finally switch on  $Q_1$ . The transistors

Figure 1



A simple circuit prevents excessive discharge of NiMH cells.

Figure 2



The load disconnects from the battery when the voltage drops below 1.1V and reconnects when the battery charges above 1.3V.

are low-threshold MOSFETs from Supertex (www.supertex.com). The circuit uses no trimming resistors. You can select  $IC_1$  and  $IC_2$  off the shelf with 100-mV steps and 2% switching-point accuracy. You can adapt the circuit for the

higher voltages of Li-ion batteries by selecting the voltage detectors.

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## Two diodes change demagnetization-signal polarity

Christophe Basso, On Semiconductor, Toulouse, Cedex, France

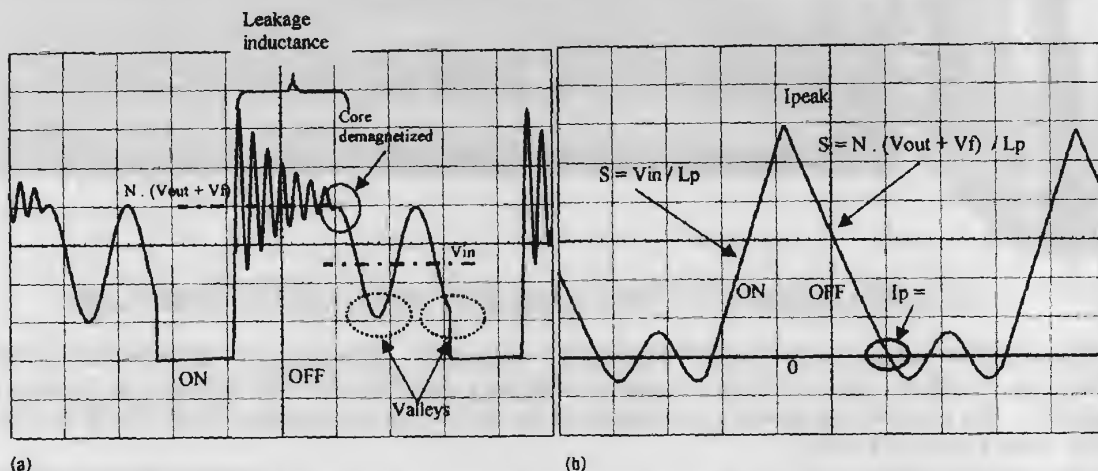
**P**OWER-SUPPLY DESIGNERS usually like flyback converters to operate in DCM (discontinuous-conduction mode) rather than in CCM (continuous-conduction mode). In DCM, the flyback converter is a first-order system at low frequencies, which eases the feedback-loop compensation. You can use a low-cost secondary rectifier, thanks to soft blocking conditions. In DCM,  $I_p$  goes to zero, and the diode stops conducting, whereas the power-switch turn-on event

in CCM forces the diode to brutally stop conducting. Also in DCM, valley switching ensures minimum switching losses that  $C_{oss}$  and all the parasitic capacitances bring.

In valley switching, or QR (quasiresonant) operation, the curve of the drain-source voltage,  $V_{DS}$ , of a typical flyback converter, shows that when the power switch closes, you observe a low level due to the  $R_{DS(ON)} \times I_p$  product (Figure 1a). At the switch opening,  $V_{DS}$  rises quickly

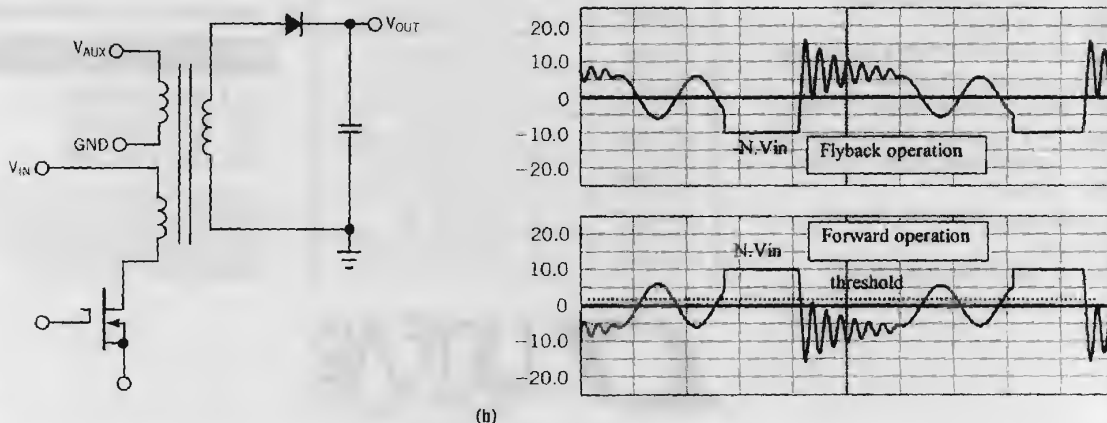
and starts to ring at a high frequency because of the leakage-inductance presence. During this time, the primary current transfers to the secondary, and a reflected level of  $N \times (V_{OUT} + V_F)$  appears on the MOSFET drain, where  $N$  is the secondary-to-primary turns ratio,  $V_{OUT}$  is the output voltage, and  $V_F$  is the secondary-diode forward drop. As soon as the primary current has fallen to zero in DCM operation, the transformer core is fully demagnetized (Figure 1b). The drain

Figure 1



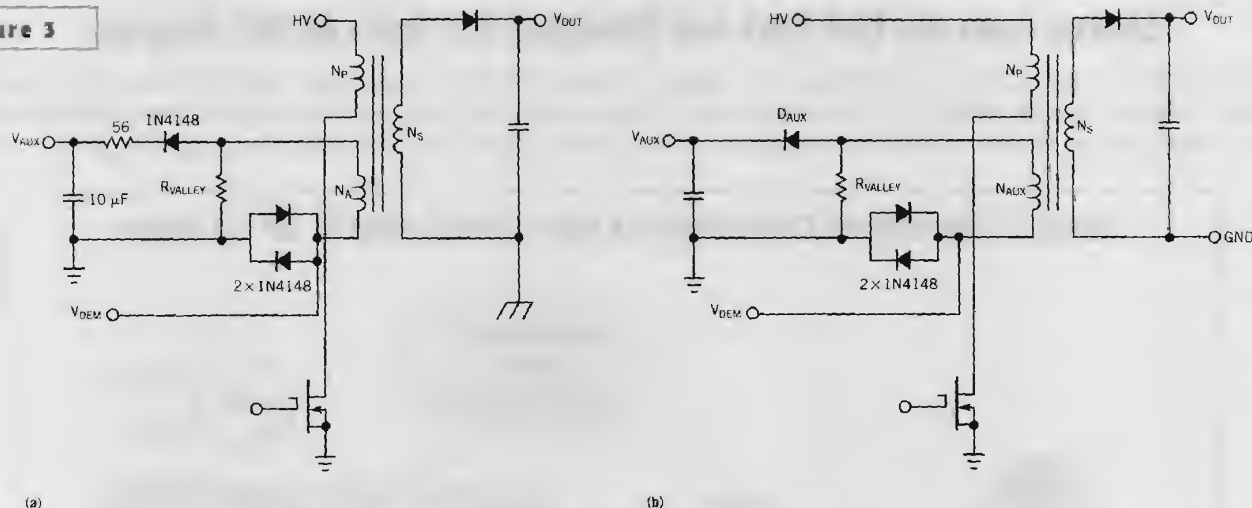
A typical drain-source waveform of a flyback converter shows high-frequency ringing (a). In DCM operation, the primary current ramps up and down to zero (b).

Figure 2



An auxiliary winding (a) lets you observe the flux image in the transformer's core for both flyback and forward operation (b).

Figure 3



A simple component arrangement allows forward-mode detection with a flyback-like PWM controller (a) or flyback-mode detection with a forward-like controller (b).

branch starts to ring but at a lower frequency than in Figure 1a because the primary inductance,  $L_p$ , is now involved.

This natural oscillation exhibits the following frequency value, where  $C_{LUMP}$  represents all of the circuit's parasitic capacitances, such as  $C_{OSS}$  and the stray capacitance from the transformer.

$$F_{RING} = \frac{1}{2\pi\sqrt{L_p \cdot C_{LUMP}}}$$

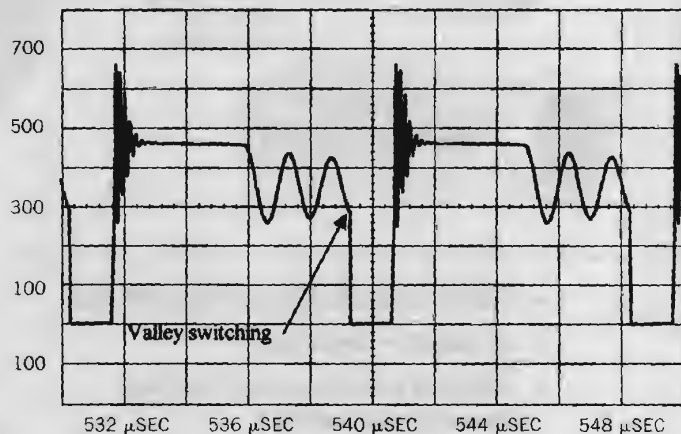
As with any sinusoidal signal, there are peaks and valleys. When you restart the switch in the valley, all the parasitic capacitance values are at their lowest possible levels. Also, the capacitive losses, which are equal to  $1/2 \times C_{LUMP} \times V_{DS}^2 \times F_{SW}$ , are small because the MOSFET is no longer the seat of turn-on losses, which removes the usual turn-on parasitics. That is the secret of QR operation.

You can easily observe the core flux through an auxiliary winding (Figure 2a). Thanks to the coupling between the windings, the auxiliary section delivers a voltage image of the core's flux through the following formula:

$$V_{AUX} = N \cdot \frac{d\phi}{dt}$$

Now, you can wire the winding either

Figure 4



When you properly adjust the time constant using  $R_{VALLEY}$ , the switch restarts in the middle of the valley.

in flyback operation, as the power winding, or in forward operation. The observed signals look the same but have different polarity (Figure 2b). Note that both signals center about ground. The problem lies in the fact that most PWM controllers accept only the flyback polarity. Typical examples include the MC33364 and MC44608 (www.onsemi.com). In battery-charger applications, you usually wire the auxiliary winding—the one that self-supplies the controller and gives the demagnetization signal—in forward mode. The reason is simple:

When the battery you charge is close to 0V, the auxiliary windings are also nearly 0V because both windings are coupled in flyback mode. By operating in forward mode, whatever happens on the secondary side is invisible, and the voltage is always there to supply the controller. However, the demagnetization signal now has the wrong polarity, and the controller doesn't restart at the core's reset event.

Figure 3a shows a way around this problem. You still wire the winding for forward operation, but you add two extra diodes in series with the winding. At

the switch closing, you apply  $N \times V_{HV}$ , where  $N$  is the ratio between the auxiliary winding,  $N_A$ , and the primary winding,  $N_P$ . You clamp  $V_{DEM}$  to  $-0.6V$ , and the current circulates through  $R_{VALLEY}$ . At the switch opening, the voltage reverses and becomes positive but clamped to  $0.6V$  on  $V_{DEM}$ . When this level collapses, the PWM controller reactivates the power switch.

You can implement this same type of circuit for PWM controllers that need a forward demagnetization signal but for which you would like to operate the aux-

iliary winding in flyback mode (Figure 3b). The problem and the cure are similar.

When you properly select  $R_{VALLEY}$ , this resistance naturally combines with sense-pin internal capacitance to add switch delay right in the middle of the wave (Figure 4).

Some controllers exhibit different demagnetization threshold levels. The MC33364 starts at around  $4V$ , and the MC44608 toggles at  $65\text{ mV}$ . Because of the diodes, you clamp  $V_{DEM}$  between  $\pm 600\text{ mV}$ , which could not trigger the

MC33364. A small offset from the internal reference to the demagnetization pin brought by a  $150\text{-k}\Omega$  resistor and a typical  $R_{VALLEY}$  of  $10\text{ k}\Omega$  have provided good circuit operation.

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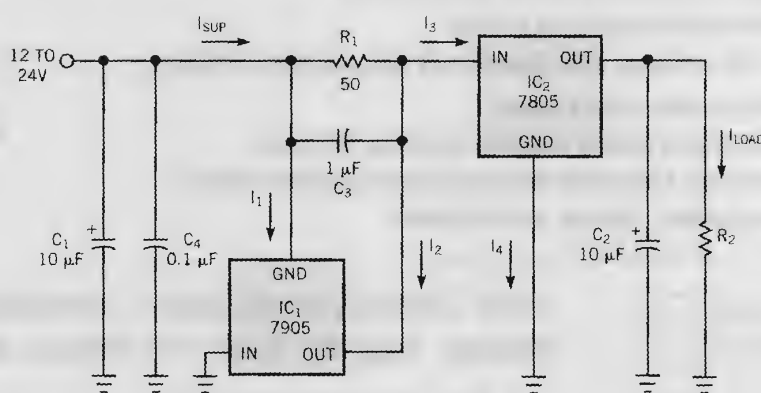
## Simple scheme keeps current drain constant

Peter Güttler, APS Software Engineering GmbH, Cologne, Germany

IT IS SOMETIMES advantageous to keep the overall current consumption of an electronic device constant. A large, seven-segment display, for example, draws nearly zero current when no segment is on to hundreds of milliamps when fully lit. This heavily varying current can cause EMI problems when a device receives its power through long cables from a remote power supply. The low-parts-count circuit in Figure 1 keeps current consumption constant.  $IC_2$  is an ordinary three-terminal regulator that supplies  $5V$  to the load,  $R_2$ .  $IC_2$  draws a total current  $I_3 = I_{LOAD} + I_4$ . ( $I_4$  is approximately  $8\text{ mA}$ , the quiescent current of  $IC_2$ .) The negative three-terminal voltage regulator,  $IC_1$ , maintains  $5V$  across  $R_1$ . The current through  $R_1$  is  $I_2 + I_3$ . So,  $I_2 = 5V/R_1 - I_3$ , and total supply current  $I_{SUP} = I_1 + 5V/R_1$ .  $I_1$  is approximately  $2\text{ mA}$ , the quiescent current of  $IC_1$ . If the load draws more current,  $IC_1$  reduces  $I_2$  and vice versa.

This regulation works well as long as  $I_3$  is smaller than  $5V/R_1$ . If the load draws more current,  $IC_1$  stops regulating and the voltage drop across  $R_1$  rises above  $5V$ . This example sets  $R_1$  at  $50\Omega$ , setting the supply current,  $I_{SUP}$ , to approximately  $102$

Figure 1



This circuit maintains a constant supply current of approximately  $102\text{ mA}$ .

$\text{mA}$ .  $C_1$  and  $C_4$  are input-filter capacitors,  $C_2$  improves ripple rejection, and  $C_3$  provides stability. Note that  $R_1$  dissipates  $(5V)^2/R_1$  and must have an adequate power rating.  $IC_1$  and  $IC_2$  may require heat-sinking. The minimum supply voltage for this circuit is  $12V$ . (The minimum input voltage for  $IC_2 = 7V + IC_2$ 's refer-

ence voltage.) If your application cannot tolerate the  $5V$  drop across  $R_1$ , try using an LM337 with a  $1.25V$  reference voltage for  $IC_1$ .

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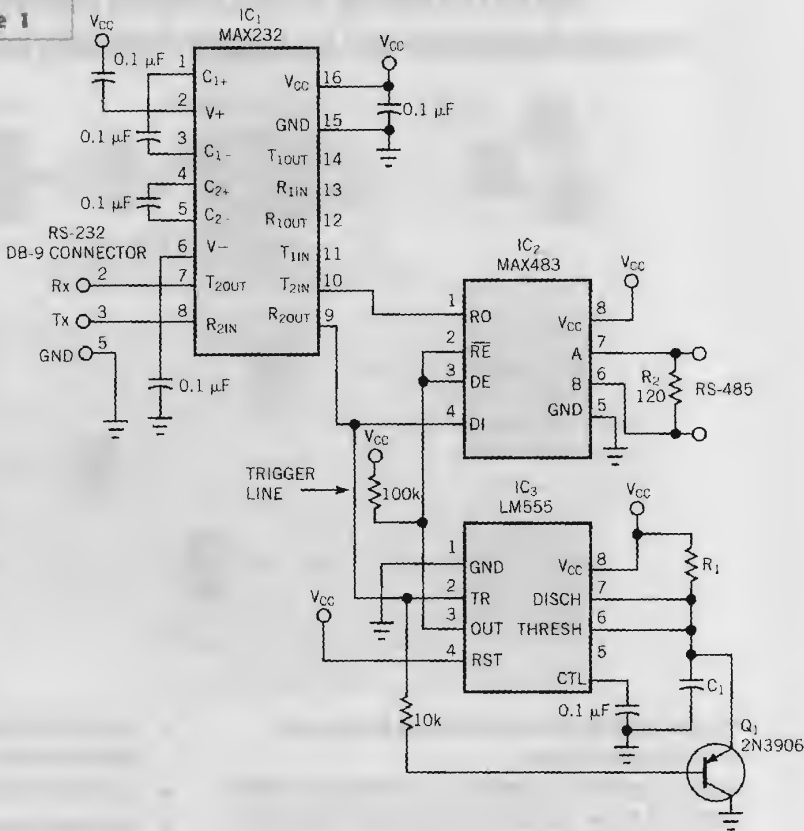
# RS-232/485 converter has automatic flow control

John Howard, Kw Aware, Ventura, CA

**R**S-485 COMMUNICATIONS can provide longer range and better noise immunity than RS-232, as well as multidrop capability. Because it does not have separate transmit and receive lines, RS-485 requires flow control. RS-232/485 converters often use one of the RS-232 handshaking lines to control direction, but several communications-software packages do not support flow control. The circuit in **Figure 1** is an RS-232/485 converter that uses the transmitted signal itself to control the flow. The circuit uses MAX232 and MAX483 interface circuits, IC<sub>1</sub> and IC<sub>2</sub> from Maxim Integrated Products ([www.maxim-ic.com](http://www.maxim-ic.com)) to convert between the ICs' respective signal levels and logic levels. Because both ICs invert the signal, the circuit preserves the original sense of the signal. The MAX483 is normally in the receiving mode. When transmission begins, the signal triggers IC<sub>3</sub>, the LM555 timer, which in turn toggles IC<sub>2</sub>'s DE and RE lines, putting the chip into the transmitting mode.

Q<sub>1</sub>, the 2N3906, fully discharges C<sub>1</sub> each time the trigger line goes low, restarting the timing cycle. The values of R<sub>1</sub> and C<sub>1</sub> determine how long IC<sub>3</sub> maintains the transmitting mode after transmission ends. This interval should be long enough such that the converter doesn't switch directions while sending characters containing long sequences of zeros. On the other hand, it shouldn't be

**Figure 1**

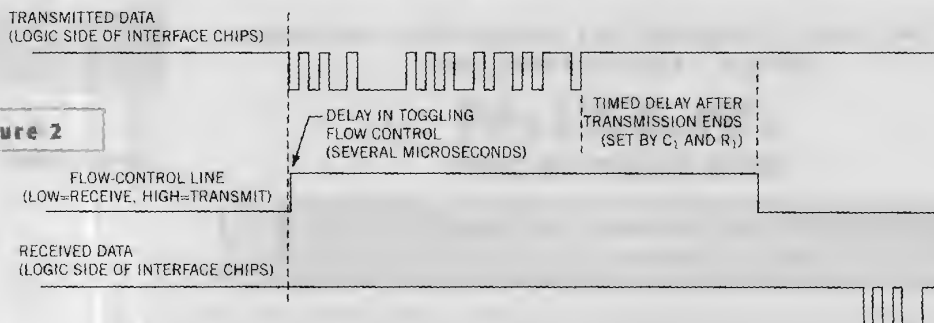


**Automatic flow control makes RS-232/485 conversion easy.**

so long that the converter misses received characters. The interval T in seconds is  $T = 1/R_1 C_1$ , where R<sub>1</sub> is in ohms, and C<sub>1</sub>

is in farads. The flow control responds within a few microseconds after transmission commences, so the converter does not miss any bits at low and medium data rates. The application for this circuit operates at 14,400 bps. **Figure 2** shows the timing of the serial and flow-control lines. The entire circuit can fit into a DB-25 (or even a DB-9) back shell.

**Figure 2**



R<sub>1</sub> and C<sub>1</sub> determine how long the transmitting mode lasts.

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# Circuit provides accurate RTD measurements

Tito Smailagich, ENIC, Belgrade, Yugoslavia

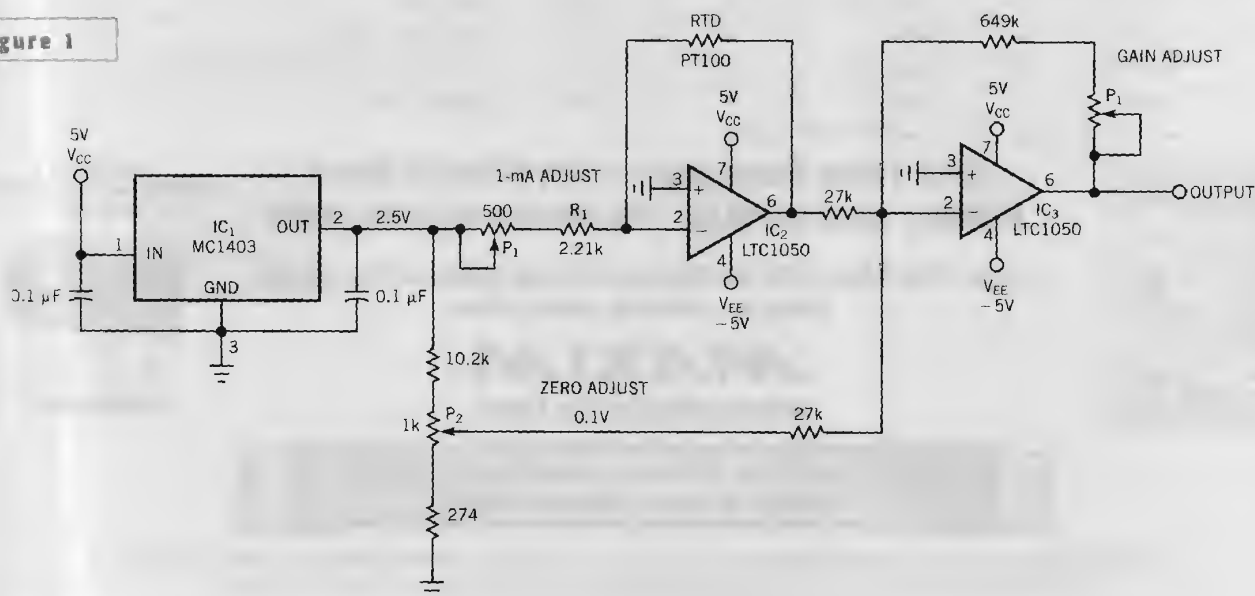
**T**HE CIRCUIT IN **Figure 1** is an efficient measuring circuit for PT100 RTD elements. IC<sub>1</sub> provides an accurate 2.5V output and, together with P<sub>1</sub> and R<sub>1</sub>, also provides a stable 1-mA current to the RTD element. The output of IC<sub>2</sub> is -0.1V. P<sub>2</sub> provides a zero adjustment (at 0°C) for IC<sub>3</sub>, an amplifier with a gain of 25. P<sub>3</sub> provides a gain adjustment. If, for example, you replace the RTD element by a fixed resistance of 124.78Ω (the RTD's resistance in **Table 1** at 64°C), you would trim P<sub>3</sub> to obtain 0.64V output. Tolerance values for Class A and B elements are ±0.35 and ±0.8°C, respectively. You can use the standard values from **Table 1** or, if you need more accuracy, you can calibrate the PT100 element in a controlled-temperature environment.

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**TABLE 1—RESISTANCE VERSUS TEMPERATURE FOR PT100 RTD ELEMENT**

°C	Ω	°C	Ω	°C	Ω	°C	Ω	°C	Ω
-20	92.16	6	102.34	32	112.45	58	122.47	84	132.42
-19	92.55	7	102.73	33	112.83	59	122.86	85	132.8
-18	92.95	8	103.12	34	113.22	60	123.24	86	133.18
-17	93.34	9	103.51	35	113.61	61	123.63	87	133.57
-16	93.73	10	103.9	36	114	62	124.01	88	133.95
-15	94.12	11	104.29	37	114.38	63	124.39	89	134.33
-14	94.52	12	104.68	38	114.77	64	124.78	90	134.71
-13	94.91	13	105.07	39	115.15	65	125.16	91	135.09
-12	95.3	14	105.46	40	115.54	66	125.54	92	135.47
-11	95.69	15	105.85	41	115.93	67	125.93	93	135.85
-10	96.09	16	106.24	42	116.31	68	126.31	94	136.23
-9	96.48	17	106.63	43	116.7	69	126.69	95	136.61
-8	96.87	18	107.02	44	117.08	70	127.08	96	136.99
-7	97.26	19	107.4	45	117.47	71	127.46	97	137.37
-6	97.65	20	107.79	46	117.86	72	127.84	98	137.75
-5	98.04	21	108.18	47	118.24	73	128.22	99	138.13
-4	98.44	22	108.57	48	118.63	74	128.61	100	138.51
-3	98.83	23	108.96	49	119.01	75	128.99	101	138.88
-2	99.22	24	109.35	50	119.4	76	129.37	102	139.26
-1	99.61	25	109.73	51	119.78	77	129.75	103	139.64
0	100	26	110.12	52	120.17	78	130.13	104	140.02
1	100.39	27	110.51	53	120.55	79	130.52	105	140.4
2	100.78	28	110.9	54	120.94	80	130.9	106	140.78
3	101.17	29	111.29	55	121.32	81	131.28	107	141.16
4	101.56	30	111.67	56	121.71	82	131.66	108	141.54
5	101.95	31	112.06	57	122.09	83	132.04	109	141.91
								110	142.29

**Figure 1**



This circuit provides accurate temperature measurements using a PT100 RTD element.